Simplified elliptic lowpass filter construction using surplus 88mH inductors

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Introduction

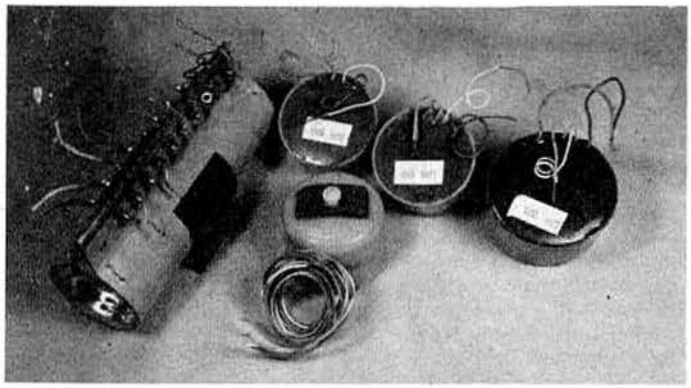
In his comprehensive article [1], G. V. Entwisle, G3MXT, explained how to apply third-method and phasing techniques in the development of ssb generators. The third-method of ssb generation required two audio-frequency lowpass filters with a sharp cutoff attenuation response, and a stopband attenuation around 45 to 50dB. G3MXT solved this lowpass filtering requirement by designing and constructing two seventh-degree elliptic filters with handwound inductors, but the large number of turns required significantly complicated the filter construction.

To simplify construction, computer-assisted calculations were made to find a suitable elliptic design that could be conveniently realized with a single unmodified stack of 88mH surplus inductors. These surplus inductor stacks are well-suited for audio filtering applications, and this article will be of interest to anyone intending to construct a third-method ssb generator, or any audio high-performance lowpass filter.

Simplified construction of modern filters

Many articles have been published on the simplified construction of modern-design passive LC filters using standard-value capacitors [2–7]. This technique is most appropriate for rf filters, where use of non-paralleled standard-value capacitors is desirable (in the rf range, the paralleling of capacitors may introduce unexpected and undesired resonances), and handwinding of inductors is easy because usually only 10 to 30 turns on a powdered-iron toroidal core gives the required inductance (in the 0·1–60μH range). For audio filtering applications, the capacitors need not have standard values because they can be paralleled in the af range without problems; however, the inductances are in the millihenry range and usually more than 200 turns are needed for each inductor. Ideally, one would use commercially manufactured toroidal inductors designed for the 300Hz to 10kHz frequency range, but the cost of these inductors is prohibitive for amateur radio applications.

The only prewound standard inductors available to the radio amateur at a reasonable price are surplus 88mH telephone line loading coils. These coils have been used by amateurs throughout the world for more than 20 years, and many articles have been published on this useful component and its properties [8–12]. Even so, it is only recently that the full capability of these toroidal inductors has been realized in the construction of relatively complex bandpass filters [13, 14]. This article explores an entirely new application of these surplus inductors in the construction of a high-performance elliptic lowpass filter, and the design and construction techniques discussed are equally applicable to the elliptic highpass filter.



The two types of surplus inductors, stack and potted. The stack dimensions are 1·5in in diameter by 4in long. The potted inductors are 1in high with diameters of 1·5 or 1·75in. One of the potted inductors is shown with a Tinnerman mounting clip. Connections are shown for 88 and 22mH

Availability of 88mH inductors

The photograph on this page shows two types of surplus 88mH inductors now available to the radio amateur. One type is an individually-potted inductor, and the other is a stack of five inductors in a cylindrical cardboard or metal case, 1·37in (35mm) diameter by 4in (102mm) long. Although these inductors are commonly referred to as 88mH inductors because this winding connection is most frequently used, they actually consist of two 22mH windings on a toroidal core of molybdenum-permalloy powder. The windings can be connected in either parallel for 22mH, or series for 88mH. This feature is important, for it provides the several combinations of inductance that are necessary to realize the elliptic design to be discussed. The inductance quadruples from 22 to 88mH when the two windings are connected in series because the turns are doubled, and since inductance varies as the square of the turns, doubling the turns quadruples the inductance (and impedance). The inductor Q is essentially identical for the parallel or series-aiding connections.

Usually the two windings are wound on opposite halves of the core to reduce interwinding capacitance, but this also results in less than perfect coupling between windings. Consequently the inductance in both parallel and series aiding connections is only about 95 per cent of what would normally be expected based on the inductance of the two separate windings. To compensate for this the separate windings are wound to give slightly more than 22mH. All windings in the parallel and series-aiding connections in this article will be assumed to have values of 22 and 88mH respectively, although the exact values may be slightly different from these nominal values.

The potted inductors shown in the photograph have their leads connected to illustrate the parallel and series aiding connections. Fig 1(a) shows the internal wiring of this inductor type and the connections for either the 22 or 88mH value. Because the inductor windings are potted in a hard compound, this inductor type can only be used in applications requiring either 22 or 88mH. Until now this constraint has made it difficult, if not impossible, to use this inductor type in complex multi-inductor filters, because turns could not be varied to change the inductance to the values required by typical filter designs. This article will demonstrate a new design technique in which a specially wired combination of five of these surplus inductors (either potted or stacked) can be used to realize the inductance values of a special seventh-degree elliptic design.

The other inductor type, perhaps more familiar to the radio amateur, is the five-inductor stack type. It is also shown in the photograph with its terminals wired to illustrate parallel and series connections. Figure 1(b) shows the internal configuration of one inductor in the stack, with directions on how to wire the terminals to get either 22 or 88mH. Because the individual inductors are not potted, they can easily be removed from the stack and modified by removing turns for any inductance value less than

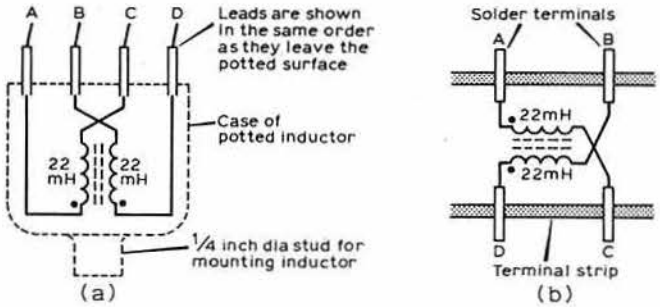


Fig 1. The internal wiring of the two coils found in the potted (a) and stack (b) inductors, and the external connections required for either the 22 or 88mH inductance values. For series-aiding connection (88mH), connect A to B to give 88mH between C and D, or connect C to D to give 88mH between A and B. For parallel-aiding connection (22mH) connect A to D and B to C to give 22mH between A and B or C and D

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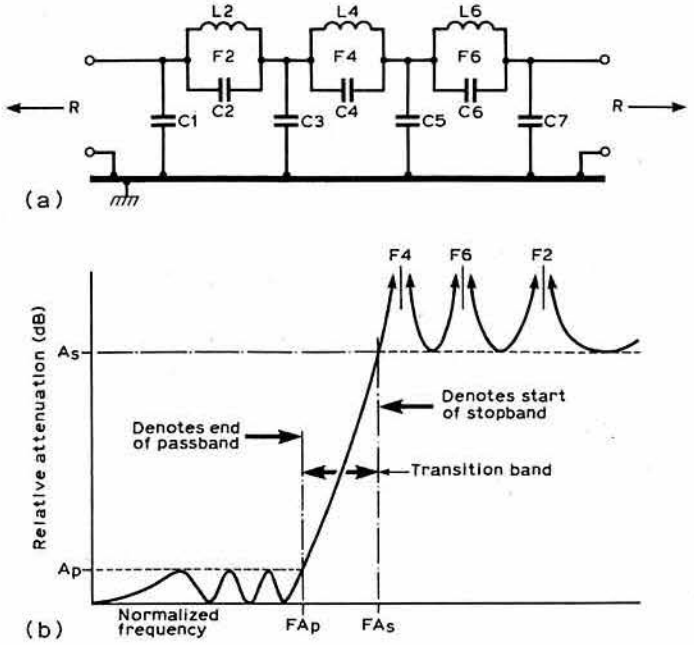


Fig 2. Elliptic lowpass filter of seventh degree: (a) schematic diagram showing component designations; (b) typical attenuation response

88mH [12]. The modified inductor can then be returned to the stack and the case resealed, or the inductor can be used separately.

In the elliptic filter application to be described, these inductors are used in their original form, so information regarding inductance versus turns removed is not required. Because of the smaller overall size and the greater ease of wiring and mounting the five-inductor stack, as compared to the potted type, the stack type can be used for building an elliptic filter suitable for installation in G3MXT's third-method circuit.

Special characteristics of the elliptic filter

An elliptic filter (as opposed to the Butterworth or Chebyshev types) was used for audio filtering in the G3MXT circuit because this type gives the most abrupt rise in attenuation. (Readers interested in learning more about this type of filter and the associated calculations should refer to [15, 16].) The seventh-degree (or three-section) elliptic design was necessary to get the required attenuation performance. Fig 2 shows the schematic diagram and the typical attenuation response of this filter.

In a typical elliptic filter design, the inductance and capacitor values usually do not repeat; that is, all the inductance values are usually different, as are the capacitor values. In comparison, the Chebyshev family of filters (the Butterworth is a special case of the Chebyshev filter in which the passband ripple amplitude is zero) has capacitor and inductor values that repeat. For example, referring to Fig 2(a), if the resonating capacitors C2, C4 and C6 are removed, the configuration is that of a seventh-degree Chebyshev filter, and for the Chebyshev values, C1 = C7, C3 = C5 and L2 = L6. This convenient relationship of the Chebyshev component values simplifies construction of the filter, but unfortunately this relationship is not available in the elliptic filter family. However, if one examines some references containing normalized elliptic designs [17, 18], it becomes obvious that there may be a special family of elliptic designs in which *L4 and L6 have identical values.* If all the elliptic designs having this unique characteristic could be found, it might be possible to find a few special designs that could be used to match the few inductor combinations available in the five-inductor stack.

After many Basic-programmed computer calculations, it was found that this unique family of elliptic designs (where L4 = L6) has a range of designs which are ideally suited for use in the audio frequency range where the minimum stopband must be greater than 45dB and the F_{A_s}/F_{A_p} ratio must be between 1·2 and 1·4. (The F_{A_s}/F_{A_p} ratio indicates the sharpness of the attenuation rise—the smaller the ratio the more abrupt the rise in attenuation.) When L4 and L6 are equal, L2 is different and larger, and the value of L2 can be expressed as a ratio relative to L4 or L6. Thus, the inductance ratios can be expressed as L2/L6 : L4/L6 : L6/L6 or "X":1:1. The computer calculations show that the L2/L6 ratio can vary between 1·2 and 1·7, with corresponding minimum stopband attenuation levels of 71·5 to 37·7dB. For the purposes of this audio filtering application, the optimum range is between 45 and 60dB. Attenuation levels greater than 60dB are generally unnecessary for audio filtering requirements, while levels less than 45dB may be too low.

Fig 3 shows the curve depicting this unique family of elliptic designs in which L4 and L6 are equal. The optimum range is indicated by the shaded portion of the curve. The normalized stopband frequency, F_{A_s}, appears on the righthand Y axis, and the corresponding minimum stopband attenuation, A_s, appears on the lefthand Y axis. It is obvious from the graph that as the stopband attenuation increases, the abruptness of the attenuation rise becomes more gradual (the normalized F_{A_s} becomes greater) for this special condition where L4 = L6.

Fortunately there are several points on the curve where the L2/L6 ratio is identical with an inductance ratio that can be obtained with a single unmodified 88mH five-inductor stack. One ratio is 132:88:88mH or 1·5:1:1. The A_s and F_{A_s} values associated with this inductance ratio are unusually well-suited for use in the G3MXT audio filtering application. That is, the attenuation rise is adequately abrupt and the minimum stopband attenuation (45·5dB) is adequate. There are also other elliptic design ratios of L2/L4 that are suitable for use with the inductor stack, but only the 1·5 ratio is of interest here. The normalized component values associated with this unique elliptic filter design were calculated from a proprietary Basic program provided by Philip Geffe, and these values will be used to find the actual component values of a filter suitable for use in the G3MXT circuit. (Mr Geffe is an internationally recognized author and authority on computer-assisted filter design and network synthesis techniques. His famous "little blue book" [19] was used by many engineers and technicians as the first understandable and practical introduction to modern filter design procedures.)

Calculation of component values from the normalized values

Table 1 shows the normalized component and frequency values of the elliptic design considered to be optimum for application in G3MXT's third-method ssb generator. An example of the calculations used to obtain these component and frequency values will demonstrate the special scaling procedure that is used when the inductor values are already known. This same procedure can be used for designing other filters having different cutoff frequencies, *so these normalized elliptic design data are universal in the sense that they can be applied for any lowpass or highpass filtering application where a single inductor stack is to be used in the filter construction.*

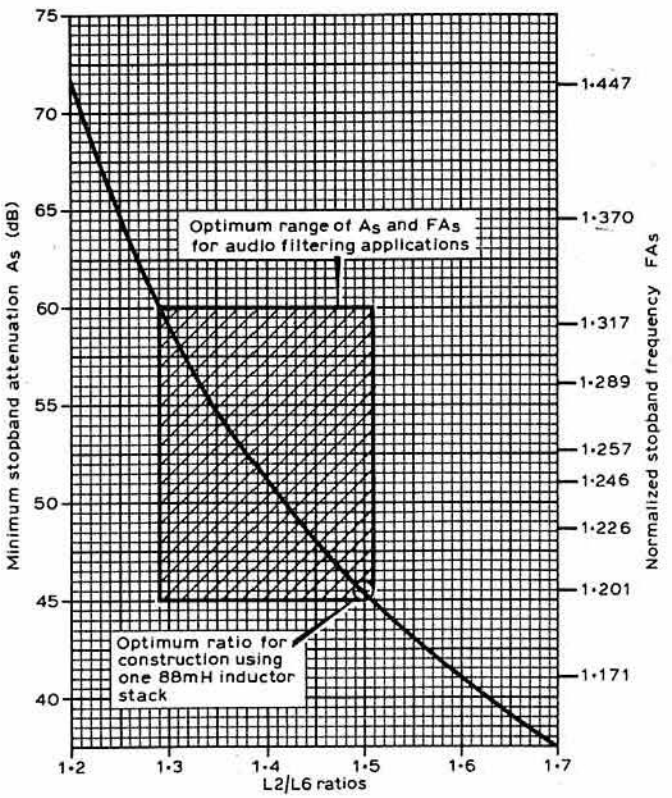


Fig 3. Stopband frequency and attenuation versus L2/L6 ratios for seventh-degree elliptic filter where L4 = L6

Table 1. Normalized parameters of an elliptic lowpass filter design for an L2/L6 ratio of 1:500

Note: This particular design is optimum for audio filtering applications when an abrupt rise in attenuation is needed and a minimum stopband attenuation of 45dB is adequate. All inductance values can be realized with a single 88mH five-inductor stack.

Termination resistance = 1·000	C1 = 0·81022F
F_{Ap} = 1·000	C2 = 0·16091F
F_{As} = 1·2012	C3 = 1·3097F
F2 = 2·2325	C4 = 0·81120F
F4 = 1·2178	C5 = 1·1427F
F6 = 1·3951	C6 = 0·61804F
A_p = 0·02831dB	C7 = 0·50758F
A_s = 45·56dB	L2 = 1·247H
Reflection coefficient = 8·060%	L4 = 0·8313H
	L6 = 0·8313H

The above element values are normalized for a cut-off frequency of 1 rad/s and 1Ω/1Ω terminations.

A typical application of an audio highpass filter is in a speech amplifier where it is desirable to attenuate voice frequencies below 300Hz. Speech frequencies below 300Hz contain little intelligence, but much power compared to the frequencies between 300 and 3,000Hz; thus a 300Hz highpass filter makes it possible to reduce the rf amplifier power load while still maintaining good speech clarity.

The usual procedure in determining the final component values of a normalized lowpass filter design (such as listed in Table 1) consists of first selecting the desired cutoff frequency (in hertz) and termination resistance (in ohms). These two parameters are used to calculate the inductance and capacitance scaling factors. The normalized component values are then scaled to the final values by multiplying the normalized values by the scaling factors. The normalized frequencies are scaled to the final values by multiplying them by the selected cutoff frequency. However, in this particular application the inductance values of L2, L4 and L6 have already been selected to be 132, 88 and 88mH; consequently, only the cutoff frequency or the termination impedance can be independently selected. Being able to independently select only one of these parameters is a compromise that must be accepted in order that the inductance values can be preselected. Because the cutoff frequency is of primary importance in this filtering application, its value will be specified, and the termination impedance will be left to fall where it may. It will be seen that this will not be a problem, as the termination impedance is such that the circuits to which the filter connects can be easily modified to provide the exact (within ±5 per cent) termination impedance required by the filter.

In the G3MXT application, it is important that maximum attenuation occurs at 1,800Hz; consequently, the scaled frequency of F4 (see Fig 2(b)) must be 1,800Hz. Since the normalized value of F4 is 1·2178 (see Table 1), then the F_{Ap} cutoff frequency must be 1,800/1·2178 or 1,478Hz. The termination impedance, R, is calculated from the selected values of L2 and F_{Ap} as follows:

$R_{(ohms)} = \omega_{Ap} \times L'2/L2 = (9{,}286{\cdot}55)(0{\cdot}132)/1{\cdot}247 = 983{\cdot}0$ where $\omega_{Ap} = 2\pi \times F_{Ap} = 9{,}286{\cdot}55$ and F_{Ap} is the cutoff frequency (1,478) in hertz, L'2 is the final scaled value in henries, and L2 is the normalized value in henries from Table 1.

The inductance and capacitance scaling factors are now calculated in the customary manner:

$$L_s = R/\omega_{Ap} = 983/9{,}286{\cdot}55 = 105{\cdot}85 \times 10^{-3}$$
$$C_s = 1/(R \times \omega_{Ap}) = 1/(983 \times 9{,}286{\cdot}55) = 0{\cdot}10954 \times 10^{-6}$$

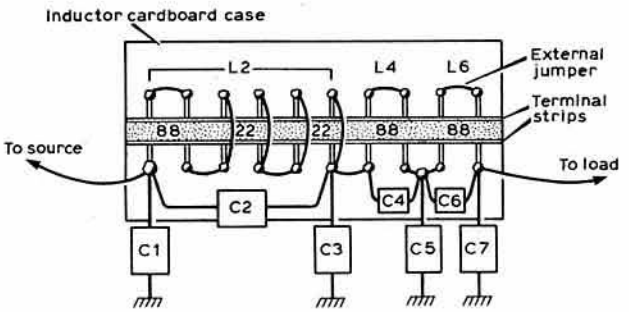
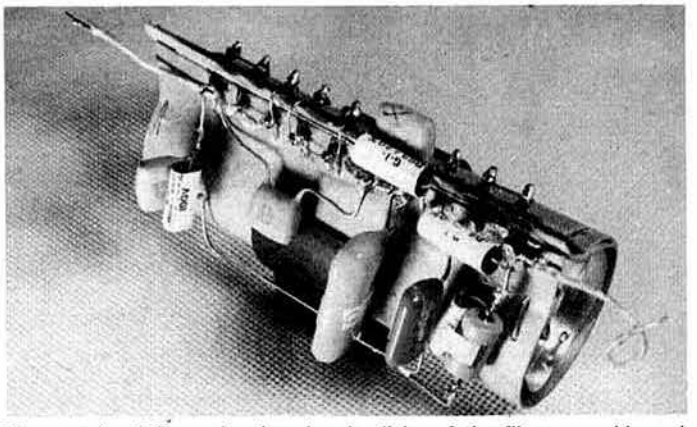


Fig 4. This shows how the jumper wires and capacitors are connected to the stack terminals. Capacitors may have to be paralleled to get the required capacitance. Component values: C1, 0·0888μF; C2, 0·0176μF; C3, 0·143μF; C4, 0·0889μF; C5, 0·125μF; C6, 0·0677μF; C7, 0·0556μF; L2, 132mH; L4, L6, 88mH. All capacitors are Mylar ±10 per cent, selected to within ±1 per cent of the design value



The completed filter, showing the simplicity of the filter assembly and mounting. C1 is at the far end of the stack. A common capacitor ground bus was used for expediency, but for maximum stopband attenuation the four capacitor ground leads must be independent of each other

The inductance values of L2, L4 and L6 are calculated in the customary manner, and if they are 132, 88 and 88mH as expected, the calculated value of R = 983Ω will be confirmed. In the following scaling equations, the L' and C' values are the final scaled values, and the L and C values are the normalized values from Table 1.

$$L'2 = L2 \times L_s = 1{\cdot}247\text{H} \times 105{\cdot}85 \times 10^{-3} = 132{\cdot}0\text{mH}$$
$$L'4 = L'6 = L4 \times L_s = 0{\cdot}8313\text{H} \times 105{\cdot}85 \times 10^{-3} = 88{\cdot}0\text{mH}$$

In a similar manner the capacitor values are calculated by multiplying each normalized C value by the capacitance scaling factor:

$$C'1 = 0{\cdot}81022\text{F}\ (0{\cdot}10954 \times 10^{-6}) = 0{\cdot}08875\mu\text{F}$$
$$C'2 = 0{\cdot}16091\text{F}\ (0{\cdot}10954 \times 10^{-6}) = 0{\cdot}01763\mu\text{F}$$
$$C'3 = 1{\cdot}3097\text{F}\ (0{\cdot}10954 \times 10^{-6}) = 0{\cdot}14346\mu\text{F}$$
$$C'4 = 0{\cdot}81120\text{F}\ (0{\cdot}10954 \times 10^{-6}) = 0{\cdot}08886\mu\text{F}$$
$$C'5 = 1{\cdot}1427\text{F}\ (0{\cdot}10954 \times 10^{-6}) = 0{\cdot}12517\mu\text{F}$$
$$C'6 = 0{\cdot}61804\text{F}\ (0{\cdot}10954 \times 10^{-6}) = 0{\cdot}06770\mu\text{F}$$
$$C'7 = 0{\cdot}50758\text{F}\ (0{\cdot}10954 \times 10^{-6}) = 0{\cdot}05560\mu\text{F}$$
$$F'_{As} = F_{As}(F_{Ap}) = 1{\cdot}2012 \times 1{,}478\text{Hz} = 1{,}775\text{Hz}$$
$$F'2 = F2\ (F_{Ap}) = 2{\cdot}2325 \times 1{,}478\text{Hz} = 3{,}300\text{Hz}$$
$$F'4 = F4\ (F_{Ap}) = 1{\cdot}2178 \times 1{,}478\text{Hz} = 1{,}800\text{Hz}$$
$$F'6 = F6\ (F_{Ap}) = 1{\cdot}3951 \times 1{,}478\text{Hz} = 2{,}062\text{Hz}$$

The scaled frequencies are found by multiplying each normalized F-value by the F_{Ap} cutoff frequency of 1,478Hz.

Filter assembly

The values of shunt capacitors C1, C3, C5 and C7 are relatively non-critical, and a capacitance tolerance of five per cent will be satisfactory. In comparison, the values of C2, C4 and C6 are critical because they determine the resonant frequencies of F2, F4 and F6. Because these resonant frequencies are primarily responsible for providing the proper stopband response, the three tuned circuits should be carefully adjusted for the design frequencies instead of using capacitors with the exact design values. Even if the exact design capacitance values are used, correct tuning of the resonant circuits is not assured because the inductor values are only nominal, and they may vary as much as two per cent.

Fig 4 shows a suggested method of wiring the five-inductor stack. A suitable assembly and tuning procedure is to connect all filter components in accordance with Fig 4, except for C2, C4 and C6. Place the assembly between a signal generator and a resistive load having the impedance values required by the design. Set the generator frequency to F2 using a frequency counter, and monitor the load voltage with an ac vtvm. Place the value of capacitance across L2 that produces resonance within about 0·2 per cent of the design value of F2. Resonance will be indicated by a null of 55dB or more below the maximum load voltage. Repeat this procedure for F6 and F4 to ensure that all three resonant circuits are correctly tuned. The actual capacitance values used should be within a few per cent of the calculated values.

For best stopband attenuation performance, the ground connections of capacitors C1, C3, C5 and C7 should be via separate leads to the chassis ground connection. The alternate grounding method of using a single common ground lead between the capacitors and the chassis ground should be avoided because the common impedance of this lead will provide a path for stopband signals to bypass the filter, thus causing the maximum stopband attenuation peaks to be less than they should be.

A photograph of the lowpass filter wired mostly in accordance with Fig 4 is shown on this page. For expediency, a common ground bus was used,

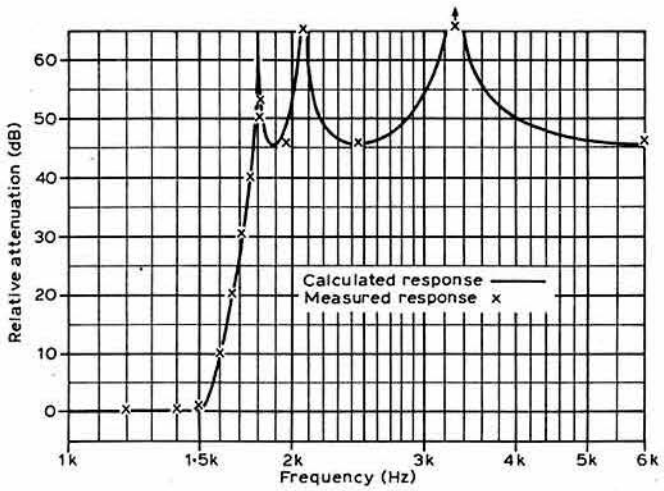


Fig 5. Computer-calculated and measured attenuation responses of elliptic lowpass filter. L4 = L6 and L2/L4 = 1·500 for R = 983Ω, F_{Ap} = 1,478Hz, RC = 8·06 per cent, A_p = 0·0283dB and A_s = 45·6dB.

Measured attenuation peak at 1,810Hz = 53dB
Measured attenuation peak at 2,062Hz = 65dB
Measured attenuation peak at 3,304Hz = 75dB
Measured insertion loss = 0·4dB at 1kHz

but this is not advisable as previously explained. Many polyester-type capacitors having nominal values of 0·15 and 0·12μF were measured to an accuracy of 0·1 per cent using a digital capacitance meter. From this selection, single capacitors were selected for C3 and C5 within five per cent of the design values. The remaining values of capacitance were obtained by paralleling capacitors as required. Although C7 is near the nominal value of 0·056μF, there were no capacitors of this nominal value on hand, so C7 was made from paralleled 0·01 and 0·047μF capacitors which measured on the low side of the standard values.

The entire filter assembly is conveniently mounted on a chassis with a standard plastic mounting clip designed to hold a 1·37in (35mm) diameter component. The convenience of this filter design, assembly and mounting technique is obvious from the photograph of the completed filter.

Filter performance

Fig 5 shows the measured and calculated relative attenuation responses of the elliptic lowpass filter recommended for use in G3MXT's circuit. A computer-controlled plotter was used to plot the attenuation levels after they were computer-calculated using a proprietary Basic network analysis program provided by Philip Geffe. The construction of the filter was started only after the calculated response curve indicated that the design was correct. This could be determined by noting that the calculated response curve agreed with the F_{A_p} cutoff frequency of 1,478Hz and with the F4, F6 and F2 peak attenuation frequencies. In addition, the plotted A_s level agreed with the expected 45·6dB value given in Table 1.

After the filter was constructed the attenuation versus frequency relative to 1kHz was measured. The measured attenuation levels are indicated by the "X"s on the graph, and they agree very closely with the computer-calculated plotted values. The insertion loss of the filter at 1kHz was an insignificant 0·4dB.

The good agreement between the measured and calculated attenuation indicates that this design should be easy to duplicate. For best results, the source and load impedances seen by the filter should be within ± 5 per cent of the design value.

Additional lowpass filter applications using the special elliptic design

By scaling the normalized component and frequency values given in Table 1 to other cutoff frequencies, this elliptic design can be applied to other lowpass filtering applications. For example, a 3kHz lowpass filter is frequently used to restrict the upper audio frequencies of voice signals before amplification or transmission and an elliptic filter can be used to provide this filtering.

Since the L'2 and L2 inductance values are fixed at 0·132 and 1·247H, an equation can be formed to show how the filter termination impedance varies with the F_{A_p} cutoff frequency. This equation is:

$$R_{(ohms)} = 2\pi(0{\cdot}132)F_{Ap}/1{\cdot}247 = (0{\cdot}6651)F_{Ap(Hz)}$$

Thus, for a cutoff frequency of 3kHz, the required filter termination impedance is 1,995·3Ω. If the cutoff frequency is slightly increased to 3,007·1Hz, the termination impedance will be an even 2,000Ω, which is preferred to demonstrate the following recommended design procedure.

For those applications where it is desired to insert the filter in an 8 or 500Ω audio system, filter impedance levels above 1,500Ω become progressively more difficult to match. This is because suitable 8Ω matching transformers in which the high-impedance windings are 1,200Ω or less are more commonly available. Consequently, for cutoff frequencies above 1,800Hz it may be advantageous to use the inductor stack in such a manner that its inductance values are one quarter of the previously discussed value. Since the filter impedance level varies directly with the inductance level, the impedance level will also be quartered. Thus the 3kHz filter can also be constructed for an impedance of 2,000/4 = 500Ω.

By a fortunate coincidence, this impedance level can be matched by the commonly available 8/500Ω matching transformer, if the filter must be used in an 8Ω audio system. In addition, because 500Ω is a commonly used audio transmission line impedance, this 3kHz filter can be placed directly in such lines. In this particular case both the filter impedance and the cutoff frequency are simultaneously optimum for many speech filtering applications.

To reduce the filter impedance level, all of the inductors in the stack are wired in the parallel aiding connection to give 22mH for each inductor. Both L4 and L6 are realized with one inductor, and the remaining three inductors are wired to give 1·5(22) = 33mH for L2. This is done by wiring two of the three inductors in parallel to give 11mH, and then connecting this combination in series with the remaining 22mH inductor to give the required 33mH. The normalized capacitor, inductance and frequency values are then scaled to their final values as previously explained. If the impedance and cutoff frequency are 500Ω and 3,007Hz, the C1 to C7 values will be 0·0858, 0·0170, 0·139, 0·0859, 0·121, 0·0654 and 0·0537μF respectively, and the L2, L4 and L6 values will be 33, 22 and 22mH. Note that because the impedance level was reduced by about half (from 983 to 500Ω) while the cutoff frequency was raised by about twice (from 1,500 to 3,007·1Hz), the old and new capacitor values are very similar because the effect of the impedance and cutoff frequency changes almost cancel each other. This effect can be anticipated from the previously-stated scaling equations for L_s and C_s. In comparison, the inductance values are quartered in the new design.

Summary

The main difficulty in constructing the typical audio filter is in realizing the inductive components. This is especially true for the radio amateur when the filter construction involves only one or two filters and the purchase of commercial inductors is not economically feasible. This article has explained how this problem can be solved by making use of a unique family of seventh degree elliptic designs in which the ratios of the three inductance values exactly match those same inductance ratios that are present in the commonly available 88mH surplus inductor stack. By properly wiring one unmodified inductor stack, it is possible to obtain the three inductor values required.

From the several possible elliptic designs that are realizable with one inductor stack, one design was selected that was optimum for the most frequently encountered audio filtering applications. A design example was explained for an application requiring a cutoff frequency of 1,500Hz, and with a minimum stopband attenuation of 46dB. Another example demonstrated how to reduce the filter impedance level to more practical levels for higher cutoff frequencies.

It is expected that this elliptic lowpass filter design will be widely used in future amateur audio filtering applications because, in combination with the surplus inductor stack, the filter is unusually easy to construct, and its attenuation performance is adequate for practically all audio filtering requirements.

How to get the 88mH surplus inductor stacks

Through the courtesy and co-operation of the Chesapeake & Potomac Telephone Company of Maryland, these inductor stacks are being made available to the author for distribution to the radio amateur fraternity. Arrangements for distributing these inductor stacks in the UK have been made with the Rev George Dobbs, G3RJV. Send a stamped self-addressed envelope to him at 17 Aspen Drive, Chelmsley Wood, Birmingham B37 7QX (Tel 021-770 5918), for further information on how to obtain these inductor stacks. Be sure to state your callsign and describe your intended application, as requests will be considered only from those having bona-fide amateur radio applications.

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does not matter how well the harmonics of f_t are suppressed in the transmitter, since the diode will create these harmonics if they are not already present.

If a communications receiver is connected to CS2 (Fig 1), and if the separation between any one of the parasitic harmonics and any one of the f_n is equal to the frequency of the receiver, the difference frequency will be received. If no parasitics are present, nothing will be heard: hence we have a parasitic detector.

At this point one may wonder where to tune the receiver. The answer is: "it does not really matter". Say that the receiver is tuned to 29MHz, usually a quiet part of the band. Somewhere in the radio frequency spectrum there will be an f_n that is separated from some p_m by a frequency difference close to 29MHz. Since the receiver passband is typically only 6kHz, it might be surmised that the difference frequency would have to be within a few kilohertz of 29MHz in order to be heard.

It must be remembered, however, that f_n is a harmonic of a crystal-controlled frequency, or of a stable vfo. On the other hand the parasitic is extremely unstable and can be moved about by any slight change, such as placing a hand near the transistor, detuning LC circuits, changing drive level or supply voltage etc. Any of these actions will cause the parasitic and its harmonics to sweep wildly about the rf spectrum, and each time one of the p_m sweeps past a point in the spectrum that is separated by 29MHz from any one of the f_n, a plop will be heard from the receiver. Usually there are a large number of plops as a finger is touched to the tuned circuits of a parasitic-producing stage. (*Fingers are not recommended where high rf or dc voltages are present*). If there are no parasitics, nothing will be heard from the speaker.

In terms of noise figure or conversion loss, the 1N914 makes a contemptible mixer, but this does not really matter as the mixer is backed up by the excellent sensitivity of a communications receiver. Even with a 40dB conversion loss the strong parasitic "signal" will inevitably be heard. Sometimes the parasitic will take the form of super-regeneration or "squegging" which will sound like a squeal or hiss in the receiver. Eventually one will learn to recognize all the different kinds of nasty noises that unstable transistors can make.

When all parasitics are completely eliminated and the rig is unconditionally stable, it will be possible to severely detune any of its LC circuits and hear nothing but background noise from the receiver.

In Fig 1 the combination of L1 and C2 forms a high L to C ratio series-tuned circuit which is resonant at 29MHz or whatever "i. f." is chosen. This prevents rf from the rig under test from being lost into the receiver, but permits the desired mixer products to pass. Of course, when testing a rig with output near 29MHz, the receiver will need to be switched to some other band, such as (say) 21MHz, and C2 adjusted accordingly. With the values shown in Fig 1, L1-C2 will resonate at about 12MHz with C2 at maximum capacitance, and at minimum capacitance the upper frequency limit is about 35MHz. C2 can be adjusted to resonate L1 to the desired frequency by running a strong steady signal from an antenna or signal generator into CS1 and adjusting for maximum S-meter reading.

For best results the receiver will be operated on a.m. (if available), maximum bandwidth, agc on, maximum rf gain, noise blanker off, and audio gain set at a comfortable level. Accidental receiver front-end damage is prevented by the limiting action of diodes D2 and D3. 1N4001 rectifier diodes prove to be better limiters at 29MHz than 1N914s and are less susceptible to burn out.

Construction

Construction details should be fairly apparent from the photographs. The device is built on a 2·5 by 4·5in (63 by 114mm) plate of 14 gauge (64mm) brass. This forms the cover for a 2·5 by 4·5 by 1in (63 by 114 by 25mm) aluminum chassis. It is important that short leads be used on the five resistors. They are arranged like spokes of a wheel around CS1 and soldered directly to its centre pin; the other ends are soldered to the brass plate.

To end up with exactly 50Ω, the five resistors should average 250Ω each. However, it is not generally realized that the resistance of composition resistors with short leads is permanently raised when they are soldered. This is caused by the heat that is conducted along the wire lead into the body of the resistor. It is best to start with presoldered values in the area of 240Ω in order to end up with an average of 250Ω after soldering: 220Ω 2W resistors are more common than 240Ω resistors, and can also be used if higher than nominal values are selected with an ohmmeter.

In the upper vhf region the diode rectification efficiency will fall off as frequency is increased. This decline can be compensated to some extent by the simple expedient of using long leads on D1. Long leads act as "peaking" inductance, and improve the high frequency response in a manner similar to the way peaking coils extend the frequency response of video amplifiers. The leads on D1 should each be about 0·5in (13mm) long for best results.

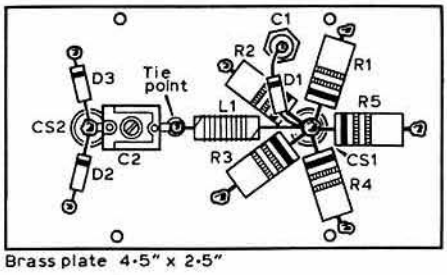

Fig 5. The five 250Ω, 2W resistors are spaced 72° apart around the input coaxial socket and are soldered directly to the brass plate with short leads

The power formula and the curve of Fig 2 will be accurate with a standard 11MΩ resistance vtvm or fet vm. Neither the curve nor the formula will be accurate for a voltmeter of different resistance. A voltmeter of more than 11MΩ input resistance can be used if it is shunted by a suitable resistor to bring the combination down to a value of 11MΩ. For instance, a 100MΩ meter would be shunted with a resistor of about 12MΩ.

If one is interested only in very low power levels, sensitivity in the milliwatt region can be improved by using a germanium diode for D1. But 20W requires a diode with a piv rating of about 90V, which is more than most germanium diodes can handle. Of course, if a germanium is used the power formula and the curve of Fig 2 will have to be modified. □

SIMPLIFIED ELLIPTIC LOW PASS FILTER CONSTRUCTION USING SURPLUS 88mH INDUCTORS

(Continued from page 321)

Acknowledgements

The author gratefully acknowledges the assistance of Mike Barge, of Honeywell, for developing the computer programs and performing the computer calculations required for this article, and of Rex Cox, of Honeywell, and Joseph Gutowski, of EWC Inc, for their review of the manuscript.

References

[1] "The G3MXT third-method Mk2 and polyphase Mk2 ssb generators", G. V. Entwisle. *Rad Com* December 1981.
[2] "Elliptic lowpass filters for transistor amplifiers", E. Wetherhold, W3NQN. *Ham Radio* January 1981.
[3] "Lowpass filters for amateur radio transmitters". *QST* December 1979.
[4] "7-element Chebyshev filters using standard-value capacitors". *RF Design* February and June 1980.
[5] "Design 7-element lowpass filters using standard-value capacitors". *EDN* Vol 26, No 1 7 January 1981.
[6] "Lowpass Chebyshev filters use standard-value capacitors". Engineer's Notebook, p160, *Electronics* 19 June 1980.
[7] "Standard C, L-input filters stabilize hf transistor amplifiers". Engineer's Notebook, p155, *Electronics* 3 November 1981.
[8] "An amateur application of modern filter design", E. Wetherhold, W3NQN. *QST* July 1966.
[9] "An RTTY bandpass filter for 1275/2125 cps". *QST* August 1967.
[10] "An RTTY bandpass filter for 2125-2975 cps". *QST* April 1968.
[11]"Inductance and Q of modified surplus toroidal inductors". *QST* September 1968.
[12] "88-mH inductors—A trap!", Harold T. Mitchell, N0ARQ. *QST* January 1983.
[13] "Modern design of a CW filter using 88- and 44-mH surplus inductors". E. Wetherhold, W3NQN. *QST* December 1980.
[14] "High-performance CW filter". *Ham Radio* April 1981.
[15] "Modern filter design for the radio amateur", W. H. Allen, MBE, G2UJ. *Rad Com* August 1971.
[16] "An introduction to elliptic filters for the radio amateur", John Wilkinson, BSc, G4HGT. *Rad Com* February 1983.
[17] *The Design of Filters Using the Catalog of Normalized Lowpass Filters,* R. Saal. Telefunken, 1966.
[18] *Handbook of Filter Synthesis,* A. Zverev. John Wiley & Sons, 1967.
[19] *Simplified Modern Filter Design,* P. Geffe. John F. Rider, 1963. □